SINGLE SIDEBAND FOR AERONAUTICAL COMMUNICATION

Edward C. Waller

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Lieutenant, United States Navy

Submitted in partial fulfillment of
the requirements for the degree of

MASTER OF SCIENCE
IN
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ABSTRACT

At the present state of development of the single sideband system of radio communication, there are presented many proposals as to how to most efficiently transmit the required signal. Most of these proposed methods have had limited application in either telephony or amateur circuits but none, as yet, has been proven by constant use in an aircraft environment.

The following thesis discusses these various proposals with a view to choosing the optimum system for a 2-30 Mcs airborne transmitter, capable of transmitting a 2700 cycle audio spectrum at any one of 28000 channels spaced one kilocycle apart. The peak power output of such a transmitter would normally be in the 500 to 1000 watt class for long range high frequency transmission.

The author wishes to express gratitude to Professor W. C. Smith of the U. S. Naval Postgraduate School for his encouragement and assistance in the preparation of this paper, and to Miss Georgia Lyke, Reference Librarian at the U. S. Naval Postgraduate School, for her kind cooperation regarding the reference work required.

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TABLE OF SYMBOLS AND ABBREVIATIONS

AM	Amplitude Modulation
db	Decibel
e_g	Instantaneous Grid Voltage
e_o	Instantaneous Output Voltage
i_b	Instantaneous Plate Current
r.f.	Radio Frequency
SSB	Single Sideband
S/N	Signal to Noise Ratio
w_c	Carrier Frequency
w_m	Signal Frequency

CHAPTER I

INTRODUCTION

The single sideband system of radio communications, developed in theory in 1915 by John R. Carson, has just begun to come into its own in 1957. Although much work has been done during and since the 1930's by the Bell System for overseas telephony and in recent years by the radio amateurs in furthering the applications of this system, it has not been until the past two years that the pressure of frequency allocation in the civilian and in the military has led to the developments which make this system widely applicable.

As of today much has been written on this subject and many different proposals have been offered for accomplishing effective single sideband communication. The scope of this thesis will be limited to a survey and examination of the various methods proposed for single sideband transmission, keeping in mind the basic needs of an airborne communication system. The author does not mean to slight the problem of reception of single sideband signals, but it is felt that the aircraft transmitter is by far the biggest weight, space, and power user and that by limiting the scope of the thesis more will be gained both by the author and the reader. The author's thinking is oriented towards the 2-30 Mcs long range airborne transmitter for this is where single sideband will have its initial application.

Before we commence the study of single sideband, it would be appropriate to define the basic needs of an airborne transmitter. These are:

1. Reliability of communication

2. Efficiency of power consumption

3. Simplicity of operation

4. Light in weight and small in size

The remainder of the introduction will be a comparison of the single sideband to the conventional amplitude modulation system it will replace in an attempt to bring out the tremendous advantage of single sideband in regard to items one, two, and four above. The remaining chapters of the thesis will evaluate various methods proposed, with a view to obtaining the optimum components for an airborne transmitter. Laboratory research on the balanced modulator was accomplished and will be described in the last chapter.

I. Single sideband versus amplitude modulation

a. Frequency spectrum

Considering a voice signal transmitted containing frequencies from 100-5000 cps, the AM system would require a full 10 kilocycle bandwidth plus the allowed 500 parts per million frequency tolerance which at 20 Mcs would amount to another 10 kilocycles. By comparison, the SSB system would have a bandwidth of 4900 cycles and due to the absolute stability requirements of 1 part per million or greater, the total spectrum necessary would be about one-fourth the AM requirements at this frequency and always less than half at any frequency.

b. Power output

It is of interest to investigate the ratio of information bearing signals to carrier power in an AM system. Many authors have shown that at 100 percent modulation, only one-fourth of the power transmitted is necessary to convey information and, therefore, on a power

basis the SSB system has a 6 db advantage over the AM system. Of more importance, I believe, is the analysis of Mr. H. S. Black who comes to the conclusion that in ordinary speech the average power of an AM system is 2518 times the average power of a single sideband[1] and, on a power basis, this is what the SSB system saves. In most airborne applications the flashover point of the aircraft antenna limits the load that can be put on it from the transmitter, and with this as a basis, the 6 db increase is applicable when one considers a given antenna, but a more realistic approach would be a combination of the above two where peak and average power are both considered.

c. Signal to noise ratio

Considering that the resistance noise is correlated over the entire bandwidth of a receiver, again, most authors agree that the reduction of bandwidth by one-half increases the S/N ratio 3 db for the SSB system. In addition to this gain, Mr. Black presents the fact that on a basis of equal peak power, the S/N advantage of SSB over AM is:[2]

$$10 \log_{10} 8\, V_p^2/S_p^2 \quad \text{where } V_p^2 \text{ is the peak power}$$

$$\text{and } S_p^2 \text{ is the peak sideband power}$$

d. Reliability of communications

Under typical long range propagation conditions the effects of multiple path transmission and selective fading are particularly deleterious to AM signals. This may result in partial cancellation of

[1] H. S. Black, Modulation Theory, pp. 131-134

[2] Ibid., pp. 137-138

the carrier signal or in phase or amplitude shift of either or both sidebands with respect to the carrier causing serious distortion and loss of intelligence. The SSB system, while still prone to these evils, is not so seriously affected due to the fewer number of frequency components of the wave and the lack of a requirement for the carrier to be in exact phase at the receiver.[1] Under conditions where severe fading has been observed successful SSB communication has been established when the AM circuit of similar sideband power was completely out of service.[2]

Mr. E. W. Pappenfus of the Collins Radio Company furnishes figures as to volume, weight, power, input, and cost between the two systems under discussion. His figures show that for a 1 kw SSB transmitter able to put out eight times as much information power as a 1 kw AM transmitter, the volume is 25% less, the weight 52% less, and the power consumption 57% less.[3] Cost figures also show the advantage of the SSB system as the output power requirements increase.

The five chapters that follow discuss the major components of a SSB transmitter with due regard for requirements two, three, and four listed above.

[1] S. Newman, Single Sideband, Suppressed Carrier Radio Communication, Appendix III, U.S.N.P.G.S. Thesis 1950

[2] J. F. Honey and D. K. Weaver, Jr., An Introduction to Single-Sideband Communications, Proc. I.R.E. 44, p. 1669, Dec., 1956

[3] E. W. Pappenfus, Power and Economics of Single Sideband, Proc. I.R.E. 44, Dec., 1956

CHAPTER II

THE BALANCED MODULATOR

The basic component for carrier suppression in a single sideband system is the balanced modulator. In its simplest form the balanced modulator is two Van der Bijl modulators connected in push-pull. See Figure 1. The grids of each tube are biased to the point of maximum parabolic curvature of their dynamic transfer characteristics, so that the plate current that flows through each tube can be approximated at:

$$i_b = a_2 e_g^2$$

From Fig. 1:

$$e_{g_1} = \cos w_c t + \cos w_m t$$

$$e_{g_2} = \cos w_c t - \cos w_m t$$

$$e_o \propto i_{b_1} - i_{b_2}$$

Assuming identical tube characteristics:

$$i_{b_1} = a_2(\cos w_c t + \cos w_m t)^2$$

$$i_{b_2} = a_2(\cos w_c t - \cos w_m t)^2$$

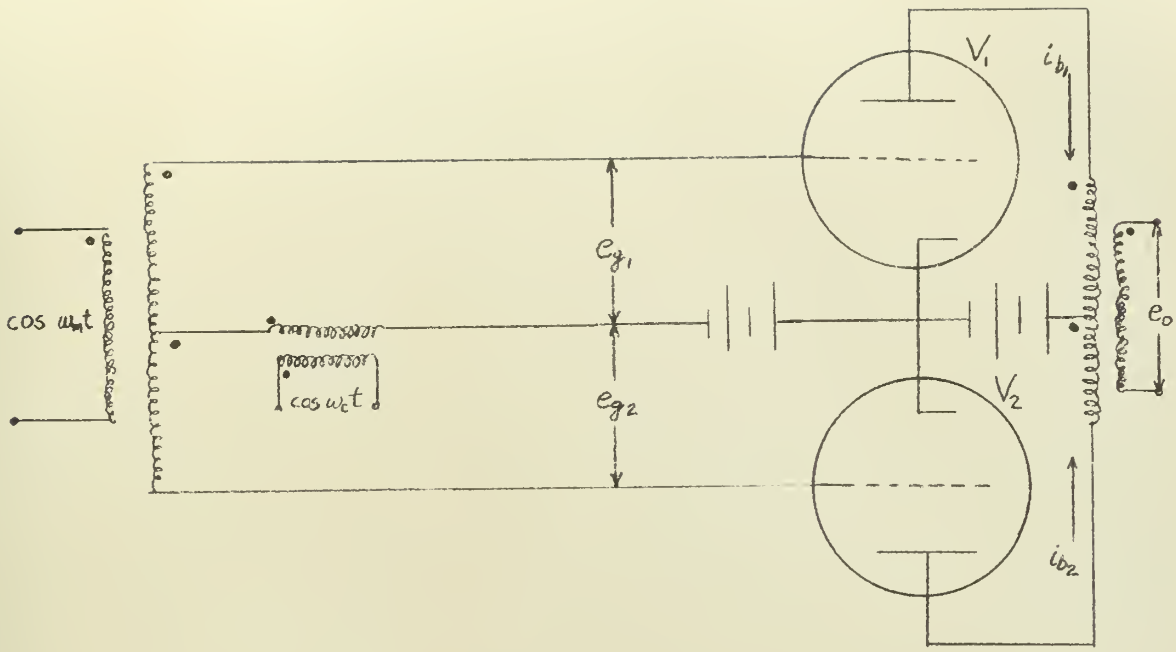
therefore, $$e_o \propto a_2 \left[(\cos w_c t + \cos w_m t)^2 - (\cos w_c t - \cos w_m t)^2\right]$$

or, $$e_o \propto a_2(4\cos w_c t \cos w_m t)$$

Trigonometric expansion shows:

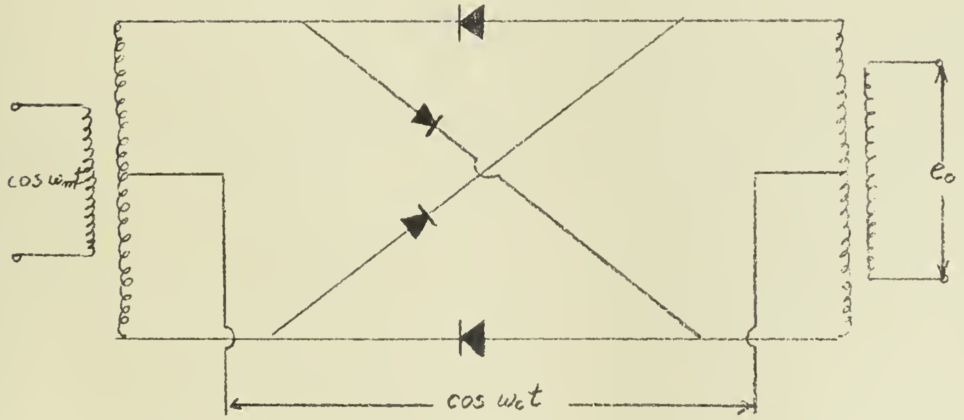
$$4a_2 \cos w_c t \cos w_m t = 2a_2 \left[\cos(w_c + w_m)t + \cos(w_c - w_m)t\right],$$

the two sidebands of the amplitude modulated signal minus the carrier.



THE BALANCED MODULATOR

FIGURE 1



DOUBLE BALANCED RING MODULATOR

FIGURE 2

The above oversimplification neglects the circuit constants. In practice, great design care must be exercised to insure that these constants are approximately the same for each path. The two dynamic transfer characteristics are never exactly identical and often show linear and higher order terms:

$$i_b = a_1 e_g + a_2 e_g^2 + a_3 e_g^3 + a_4 e_g^4 + \ldots\ldots\ldots,$$

these higher order terms creating intermodulation products, $\cos(w_c \pm 2w_m)t$, $\cos(w_c \pm 3w_m)t$, etc., that distort the wanted sidebands. The linear term allows the carrier component, $\cos w_c t$, to be a part of each plate current, but the subtraction in the output transformer acts to balance out this component. To obtain perfect balance of all even order intermodulation products and carrier terms, the modulators must be identical, and the amplitude and time delay relations for each path must be the same. In practice this can not be accomplished, so a balancing mechanism is included to balance out the carrier term to the maximum degree possible. This balancing mechanism is a device used to alter the amplitude of the respective grid signals or the dynamic transfer characteristics of the modulators. In vacuum tube circuits, a variation in plate load, grid signal bias, or change in relative amplitudes of carrier and signal will accomplish a high degree of balance, but aging of the tubes or other circuit components tend to destroy this balance.

In practice, copper oxide and germanium varistors have largely replaced the conventional vacuum tubes as balanced modulators due to their superior performance in maintaining balance for long periods of time, not to mention their low initial cost, lack of power requirements,

and essentially infinite life. A typical balanced modulator using varistors is shown in Figure 2. Mr. R. S. Caruthers shows other basic configurations in the Bell System Technical Journal of April 1939. In this circuit, as well as all others of its type, modulation results from either the reduction or reversal of the current flow between input and output signal circuits at periodic intervals as the carrier varies the varistor resistance back and forth between low and high values. Satisfactory operation of these modulators requires a large carrier voltage with respect to the signal so that bothersome intermodulation products are reduced to a negligible level.[1]

If the "perfect switch" is assumed, the expression for this switching function is simply the Fourier expansion for a square wave,

$$e_c = 4/\pi \sum_{n=0}^{\infty} \frac{(-1)^n}{2n + 1} \cos(2n + 1)w_c t$$

where the amplitude of the square wave is assumed to be unity. The input signal, $\cos w_m t$, can be multiplied by this switching function above to obtain the output signal;

$$e_o = 2/\pi \sum_{n=0}^{\infty} \frac{(-1)^n}{2n + 1} \left\{ \cos \left[(2n + 1)w_c + w_m \right] t + \cos \left[(2n + 1)w_c - w_m \right] t \right\}$$

If the carrier voltage is large with respect to the signal, the above "perfect switch" representation holds, but if the signal voltage is large enough to participate in controlling the resistance of the varistors, the power series approach would be necessary, and the current-voltage characteristic of the varistors is not so kind as to be a parabola but rather contains many terms of sizeable coefficients, such as:

[1] R. S. Caruthers, Copper Oxide Modulators in Carrier Telephone Systems, Bell System Technical Journal 18, p. 319, April, 1939.

$$i = a_1 e + a_2 e^2 + a_3 e^3 + a_4 e^4 + \ldots\ldots \quad \text{where}$$

$$e = \cos w_c t \pm \cos w_m t$$

The higher order terms produce serious distortion products in the band pass desired. For a signal to distortion ratio of 50 db the ratio of carrier to signal amplitudes should be at least 15.[1] Chapter VII shows experimental evidence to corroborate this assertion.

Even though the varistor modulator induces a loss in the circuit, usually from 5 to 9 db,[2] the stability of effective balance has been deemed worth this price by many authors. An effective balance of about 45 db carrier suppression has been obtained with well matched modulator arms.[3] A good many circuits employ copper oxide varistors for the low frequency modulators and crystal diodes for the high frequency modulators.

Laboratory research discussed in Chapter VII included the building and testing of balanced modulators employing tubes, crystal diodes, and transistors. Testing showed that, on a comparative basis, a greater signal to distortion ratio was obtainable with vacuum tubes. It was quite noticeable, however, that the vacuum tube modulator required constant attention in regard to balance each time the ratio of carrier to signal was changed. Although the crystal diode modulator showed intermodulation distortion effects at low ratios of carrier to signal, the carrier balance was comparatively independent of this ratio. The transistors gave the poorest results of all in regard to stability of balance but this was undoubtedly due to the fact that the current-voltage relations for the

[1] A. E. Kerwien, Design of Modulation Equipment for Modern Single-Sideband Transmitters, Proc. I.R.E. 40, p. 800, July, 1952.

[2] R. S. Caruthers, Copper Oxide Modulators in Carrier Telephone Systems, Bell System Technical Journal 18, p. 323, April, 1939.

[3] A. E. Kerwien, op. cit., p. 800.

two transistors used were more divergent than individual differences between other components tested.

Considering that the varistors can be used bilaterally in transceiver design, plus the inherent stability of balance, it would seem that these would be the simplest and most efficient for airborne use.

CHAPTER III

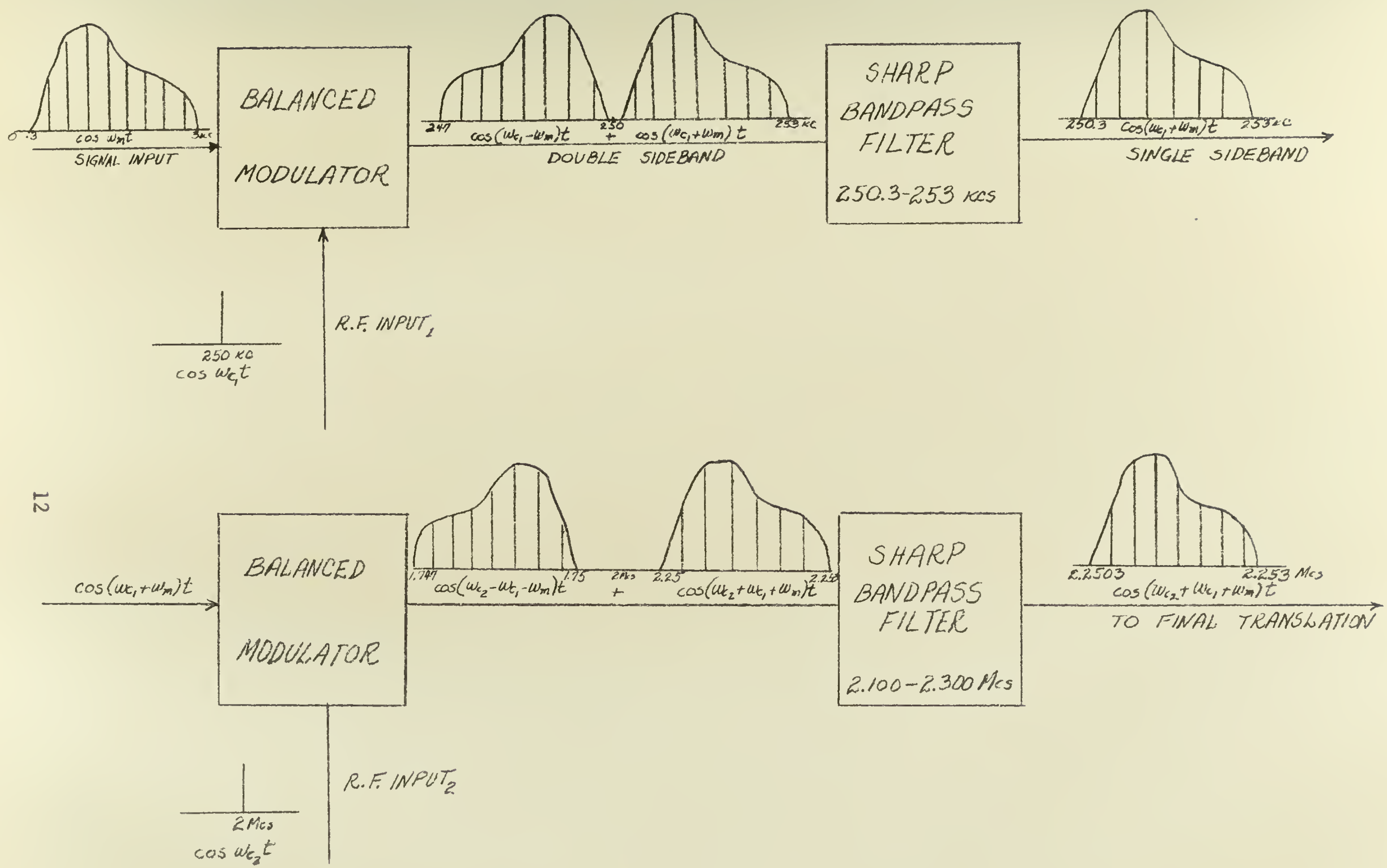
THE GENERATOR

The generator of a single sideband transmitter is the link between the input signal and the amplifiers, converting this input signal into a single sideband at the transmitting frequency. Since 1946, when Dome introduced a practical design for wideband phase shift networks,[1] there have been two basic methods employed for the generation of a single sideband signal: the filter and the phase shift methods. D. K. Weaver, Jr. has recently presented a third method[2] which may have application. These three methods will be discussed below with a view to obtaining the most practical for an airborne transmitter where simplicity and the degree of suppression of the unwanted signals are the important factors.

The filter method, shown in Figure 3, has long been the one most emphasized for variable frequency transmitters. In this method, the input signal is converted into a double sideband signal by a balanced modulator with a fixed r.f. input. A sharp bandpass filter is then employed to pass the desired signal only, rejecting all other outputs of the balanced modulator except the intermodulation products in the band pass. Due to limitations in filter design, it is necessary to use more than one frequency translation to get to final transmitter frequency in the 2-30 Mcs range. Figure 3 shows two such frequency translations employing two sharp bandpass filters. The final translation, not shown,

[1] R. B. Dome, Wide-Band Phase Shift Networks, Electronics 19, pp. 112-115, Dec., 1946.

[2] D. K. Weaver, Jr., A Third Method of Generation and Detection of Single-Sideband Signals, Proc. I.R.E. 44, pp. 1703-1705, Dec., 1956.



THE FILTER METHOD

FIGURE 3

uses the tuned amplifier stages to reject the undesired sideband from the final mixer, the frequency separation of the sidebands being more than adequate at this point. The first filter, shown in Figure 3, would be a mechanical filter and the second, a crystal lattice. Up to the era of the mechanical filter, it was necessary to use three to five frequency translations to reach final transmitting frequency when employing this method. A. E. Kerwien showed in 1952 that 125 kcs was about the maximum first translating frequency due to limitations in filter design.[1] With the advent of the electromechanical filter, it is now possible to obtain initial translation of an audio input signal at 250 kcs with 50 db of unwanted sideband rejection plus an additional 27 db of carrier rejection.[2] Even today two to three frequency translations are required to reach final transmitting frequency in the 2-30 Mcs range, but the degree of carrier and adjacent sideband suppression using this method has not yet been obtained by the other methods to be discussed. The other advantages of this method are its simplicity and use of passive elements, both of which are desirable features from a maintenance standpoint. The two main disadvantages are the need for several frequency translations and the cost and difficulty in design and manufacture of the mechanical and crystal lattice filters. Two other disadvantages, relative to the phase shift method, are its inherent inflexibility as regards choice of sideband, and the lower magnitude signal delivered to the amplifiers. The latter is due to the insertion loss of the filters, the optimum operating criteria of the

[1] A. E. Kerwien, Design of Modulation Equipment for Modern Single-Sideband Transmitters, Proc. I.R.E. 40, p. 799, July, 1952.

[2] D. L. Lundgren, Electromechanical Filters for Single-Sideband Applications, Proc. I.R.E. 44, pp. 1744-1749, Dec., 1956.

balanced modulators, and the additive relations in the phase shift method.

The phase shift method was invented by Hartley in 1928[1] but it was not until 1946 that the design of wideband phase shift networks led to wide application of this method. The concept involved in this method is one of combining the modulation products of two balanced modulators whose respective signal and r. f. inputs are ninety degrees out of phase. The proper combination of these products results in the suppression of one sideband and the reinforcement of the other. Figure 4 shows the phase shift method. The output of the upper modulator, which combines two cosine terms, contains the following products:

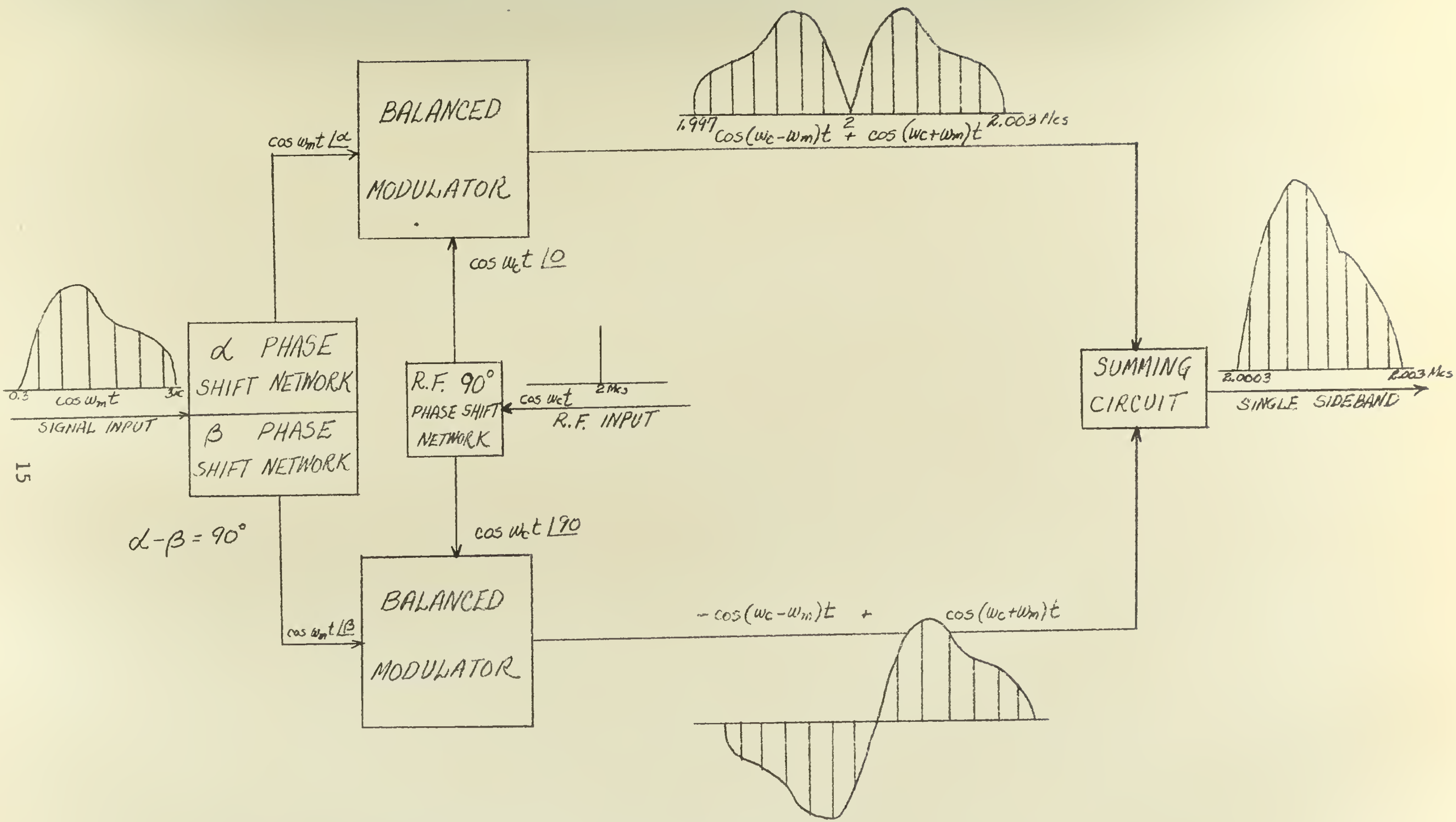
$\cos w_m t$	$\cos(w_c + w_m)t$	$\cos(w_c - w_m)t$
$\cos 3w_m t$	$\cos(2w_c + w_m)t$	$\cos(2w_c - w_m)t$
$\cos 5w_m t$	$\cos(w_c + 3w_m)t$	$\cos(w_c - 3w_m)t$
etc.,		

the first three of which are due to the linear and second order terms of the power series representation, and the latter being the unwanted intermodulation products caused by the higher order non-linearities of the transfer characteristics. These latter products are usually 40 db or more below the wanted output. The lower balanced modulator combines two sine terms with respect to the upper modulator, being that both inputs are ninety degrees phase shifted from the upper modulator's inputs. The output of this modulator contains the products:

$$-\sin w_m t \qquad \cos(w_c + w_m)t \qquad -\cos(w_c - w_m)t$$

plus similar intermodulation products as shown for the upper modulator.

[1]U. S. Patent 1,666,206.



THE PHASE SHIFT METHOD

FIGURE 4

Precise operation of these balanced modulators is necessary to insure that all unwanted products, including the linear term, are well below sideband outputs. The summing circuit reinforces the upper sideband and cancels the lower sideband as shown, or can accomplish the reverse by a simple change in polarity.

As one can imagine, in order to obtain perfect unwanted sideband suppression, amplitude and phase relations must be exact. This means that the two modulators must be identical, and the two phase shift networks must supply equal amplitude signals with exactly ninety degree phase difference to these modulators. Also, the time lag through each network and modulator must be the same. Needless to say, the above requirements are impossible to fulfill. It has been found, in practice, that a single audio phase shift network can not maintain tolerable amplitude and phase relations over the entire bandpass, but that by including a network in each signal input path, deviations have been held to 0.2° in phase and one percent in amplitude.[1] These two networks are often designated α and β networks, as shown in Figure 4, and are designed so that $\alpha - \beta$ equals ninety degrees throughout the audio frequency spectrum. It can be shown[2] that the coefficient of the undesired sideband is:

$$\sqrt{A^2 + B^2 - 2AB\cos(\Delta + \delta)}$$

where A is the amplitude of the upper modulator input signal, B is amplitude of the lower modulator input signal, Δ is the deviation from

[1] D. E. Norgaard, The Phase-Shift Method of Single-Sideband Signal Generation, Proc. I.R.E. 44, p. 1719, Dec., 1956.

[2] Ibid., p. 1720.

ninety degrees of the r.f. inputs, and δ is the deviation from ninety degrees of the signal inputs, assuming identical balanced modulators. The undesired sideband will vanish if $A = B$ and $\Delta = \delta = 0$. Through very careful design and precise balance, A can be made close enough to B, and δ sufficiently small to insure this sideband is suppressed 35-40 db. No practical r.f. phase shift network has yet been proven that will hold Δ within tolerable limits throughout a frequency range of from 2-30 Mcs. Two solutions that circumvent this problem have been offered, the more simple of the two being to use a fixed r.f. input and one final frequency translation. The second solution, presented by J. R. Hall,[1] eliminates the critically adjusted ninety degree r. f. phase shift network by use of a variable gain control in one of the signal input paths. Mr. Hall uses the fact that the above coefficient can be made zero if $A^2 + B^2 = 2AB\cos\Delta$ where Δ is different from zero, and he shows that for zero unwanted sideband output, $\tan\Delta$ is a function of A-B/A+B. By varying the modulating index of one signal with respect to the other, Mr. Hall found that the r.f. inputs could vary plus or minus thirty degrees from the desired ninety with no decrease in unwanted sideband suppression, and only 6 db decrease in desired sideband output. Employing a servomechanism loop to control the variable gain in one of the input paths, Mr. Hall's solution is surely creditable, but no doubt the increased complexity would add maintenance problems. The use of a fixed r.f. input and additional frequency translation seems simpler and should create fewer problems without the loss in signal to distortion

[1] J. R. Hall, A Phase Rotation Single-Sideband Generating System, RCA Review 16, pp. 43-51, March, 1955.

ratio that the Hall solution entails.

O. G. Villiard, a proponent of the phase shift method, spent more than a year in investigation of this method at Stanford University's Electronics Research Laboratory. His conclusions, in 1953, showed the main advantages of this method to be:[1]

1. Low cost and ease of manufacture.
2. Good frequency response to 20 kcs.
3. Ease of obtaining either sideband output desired.
4. Ability to generate the single sideband directly at final operating frequency.

The disadvantages discussed were:

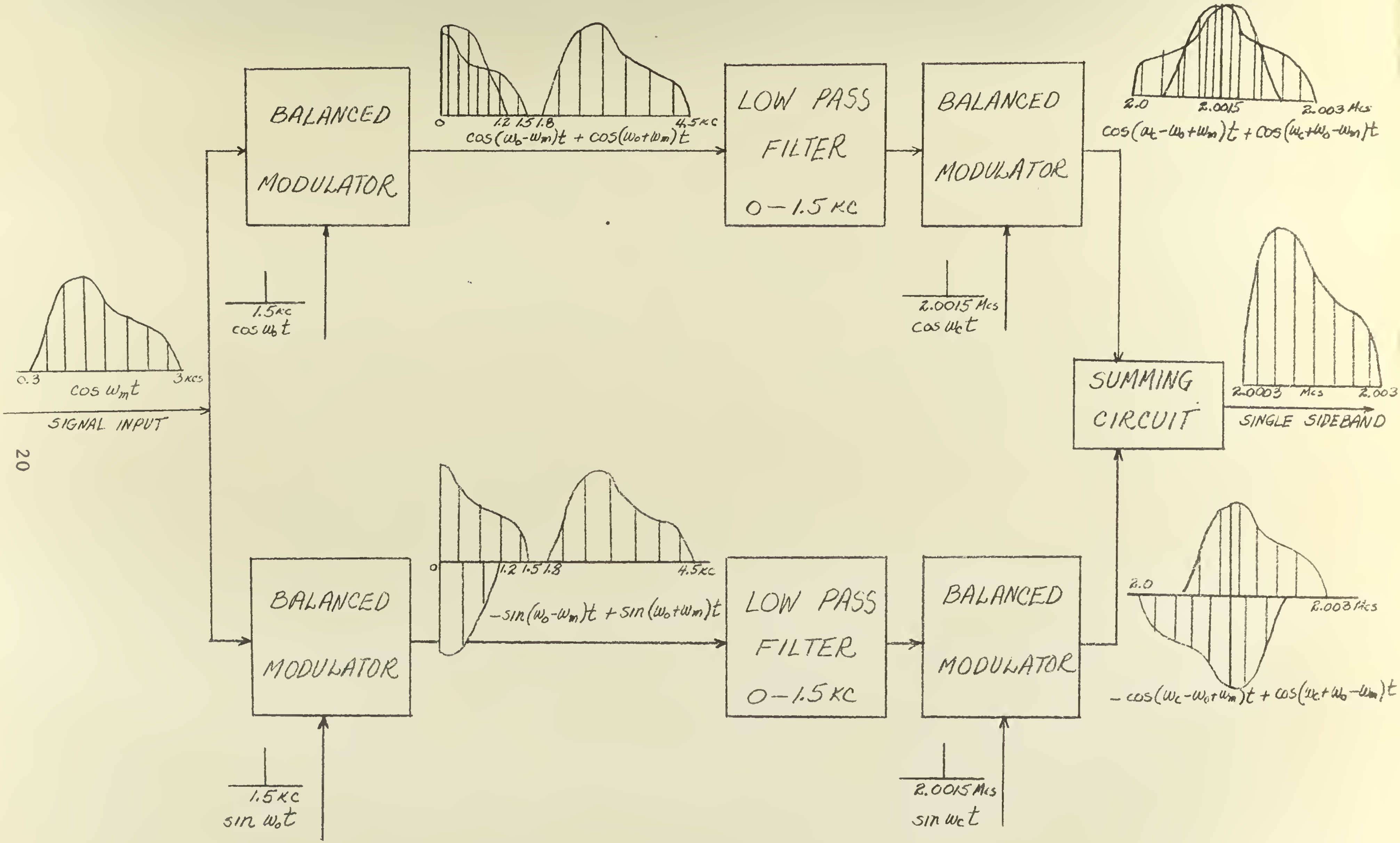
1. Phase and amplitude requirements very critical.
2. Only 40 db signal to distortion ratio in practice.
3. Need to rebalance system if variable frequency output required.

Experience has shown that the cost of the filter and the phase shift generators are about the same for tolerable signal to distortion ratio outputs. The α and β networks largely govern the system performance in a manner quite similar to the bandpass and attenuation bands of the filters, and the greater the degree of suppression desired, the greater the number of circuit branches and elements. The frequency response and ease of selecting desired sideband output are good features but not too important as far as airborne requirements are concerned. The 40 db signal to distortion ratio obtainable with this method is adequate when one considers that the amplifiers add intermodulation distortions to at least this level if not above. It should be pointed out, though, that

[1] O. G. Villiard, Communication Techniques (Modulation) Project, Final Report 1 Jun 1952 - 31 Aug 1953, Electronics Research Laboratory, Stanford University.

this is about the best obtainable signal to distortion ratio with the most complex of the networks, and that any aging of the components in the system can make this figure lower. The phase, amplitude, and time delay requirements make maintenance of this method a problem, especially in the aircraft environment. The author believes that the phase shift method is more than suited for use with a fixed frequency transmitter, but when employed for a variable frequency transmitter, does not simplify the system but rather adds to the complexity of it. For the 2-30 Mcs airborne transmitter under discussion, the phase shift method would require two balanced modulators and two complex phase shift networks. The filter method for the same transmitter would require the same number of modulators and two sharp bandpass filters. Both would require one additional frequency translation to final operating frequency.

A third method, recently introduced by D. K. Weaver, Jr., employs four balanced modulators and two conventional low pass filters feeding a summing circuit. Figure 5 shows the Weaver method. The input signal is heterodyned initially with a frequency near the center of its spectrum. The first upper modulator combines two cosine terms around the 1.5 kc shown. Recalling that the cosine of a negative angle is equal to the cosine of the same positive angle, the negative frequency terms of the lower sideband output occupy the same spectral and phase relations as the identical positive frequency terms. The upper sideband is rejected by the low pass filter and the "halved" lower sideband is heterodyned in the conventional manner as shown. With respect to the first upper modulator, the first lower modulator combines a sine term with a cosine term, the product being two sine sidebands around the 1.5 kcs shown. Trigonometric



THE WEAVER METHOD

FIGURE 5

relations show that the negative frequency terms of the lower sideband occupy the same spectrum as the identical positive term but phase shifted by 180 degrees. Again the upper sideband is rejected and the second balanced modulator combines two sine terms which expand into two cosine sidebands, the lower one being the negative inversion of the upper. As can be plainly seen, a simple summing circuit does the rest of the job as in the normal phase shift method. An additional frequency translation is required to get the signal to transmitting frequency in the 2-30 Mcs range. The circuit can be made up of simple passive elements and looks to be easy and inexpensive to manufacture. Bilateral in nature, employing crystal diodes, the uniqueness of this method is that the unwanted sideband occupies the same frequency spectrum as the desired sideband though inverted and suppressed 30 db or more. This would seem a good feature to insure the absence of adjacent channel interference when an unbalance did occur. The employment of simple inexpensive elements plus the unique spectral arrangement make this method worthy of more complete evaluation for airborne use.

The filter method, with its lack of necessity for critical adjustments in time delay, phase and amplitude relationships, and its unmatched degree of suppression of unwanted signals, is surely the most simple and easy to maintain of the methods discussed. I would, therefore, pick this method for an airborne single sideband system.

CHAPTER IV

THE OSCILLATOR

As stated previously, the single sideband system of radio communication was invented in 1915. It does seem odd that throughout the forty years that have elapsed, the tremendous advantages of this system have seen so few applications. The reason for this unfortunate fact has been the wide gap between the stability requirements for a single sideband oscillator and the inherent stability of the best of the oscillators used in practice during this period. Even today the basic limitation on the use of this system is the lack of absolute frequency stability.

In order to demodulate a single sideband signal, the receiver must resupply the carrier that was so effectively balanced and filtered out in the transmitter. It can be shown[1] that there are no absolute phase or frequency requirements for the reinserted carrier as there are for the conventional amplitude modulated signal. The stability requirement comes from the fact that reinserting the carrier at a frequency of $w_c + \delta$, an upper single sideband signal becomes $w_m - \delta$ throughout its spectrum, or a lower single sideband signal becomes $w_m + \delta$ throughout its spectrum.[2] Many authors agree that a 10-15 cps error is of no consequence and can hardly be ascertained, and all agree that a 50 cps error under good communication conditions is still intelligible but that rapid deterioration of intelligibility ensues as the error becomes more than 50 cps. Although still under study, it has been found that frequency errors in the 10-50 cps range hasten decreasing intelligibility

[1] S. Newman, Single Sideband, Suppressed Carrier Radio Communications, U.S.N. Postgraduate School Thesis, Appendix III, 1950.

[2] Ibid.

as the signal to noise ratio decreases during poor communication conditions.[1] Two other sources of frequency error, besides the instability of the oscillators, are the Doppler shift effects of the moving ionosphere and the moving aircraft. This shift is of the order of one part per million for every 670 mph relative motion directly towards or away from the r.f. source. If the ionosphere was moving up and the aircraft away at a combined rate of 670 mph with respect to a 20 Mcs carrier radiation, the frequency components of a lower single sideband signal would be shifted down 20 cps, making a natural voice sound a bit higher in pitch.

The above discussion listed the sources of error and the empirical frequency requirements for intelligible communication. What then is the stability requirement for the oscillator and how can this be obtained? Considering frequency errors of the two oscillators only, if a maximum of 10 cps is allowed at 30 Mcs, each oscillator must have a long term stability of 1.67 parts in 10^7. As the speed of aircraft is increased, the Doppler shift will assume more importance and will have to be taken care of by either automatic or manual control, such as a vernier trimmer panel control that could shift the oscillator frequency by 30-50 cps when necessary. The main problem today is in designing transmitter and receiver oscillator systems that will have the greatest stability possible.

Up to the present date, commercial single sideband applications, such as telephony, have employed a pilot carrier transmitted 10-30 db below the sideband output and have used an automatic frequency control

[1] J. F. Honey and D. K. Weaver, Jr., An Introduction to Single-Sideband Communications, Proc. I.R.E. 44, p. 1673, Dec., 1956.

circuit to maintain the receiver oscillator at the transmitter frequency. This system, while necessary in the past, loses the relative immunity to selective fading that true single sideband enjoys and is vulnerable to jamming. W. L. Firestone concluded,[1] on the basis of his studies, that modulator splatter, transmitter efficiency, peak sideband power, desensitization, and intermodulation factors were poorer with the pilot carrier than without. Today there are two systems that meet the above mentioned stability requirements: the single crystal frequency synthesis system and the stabilized master oscillator, both of which employ a standard reference oscillator. These two systems will be described below with a view to obtaining the most reliable for an airborne transmitter in the 2-30 Mcs range.

The frequency accuracy and stability of these two systems are completely dependent upon the standard reference oscillator. Using techniques now available, it is possible to produce small and rugged oscillators capable of performing in an aircraft environment with the precision equal to many laboratory standards. Besides great care in manufacture and mounting, the quartz crystal must be maintained at a constant temperature by use of an electronic oven. Craiglow and Martin[2] show long and short term stability curves for such an oscillator which, with a transistorized oven control circuit and frequency dividers, is housed in a module only 3 x 4-1/2 x 4-1/2 inches. The stability of this oscillator is shown to be greater than one part in 10^7 per month, limited

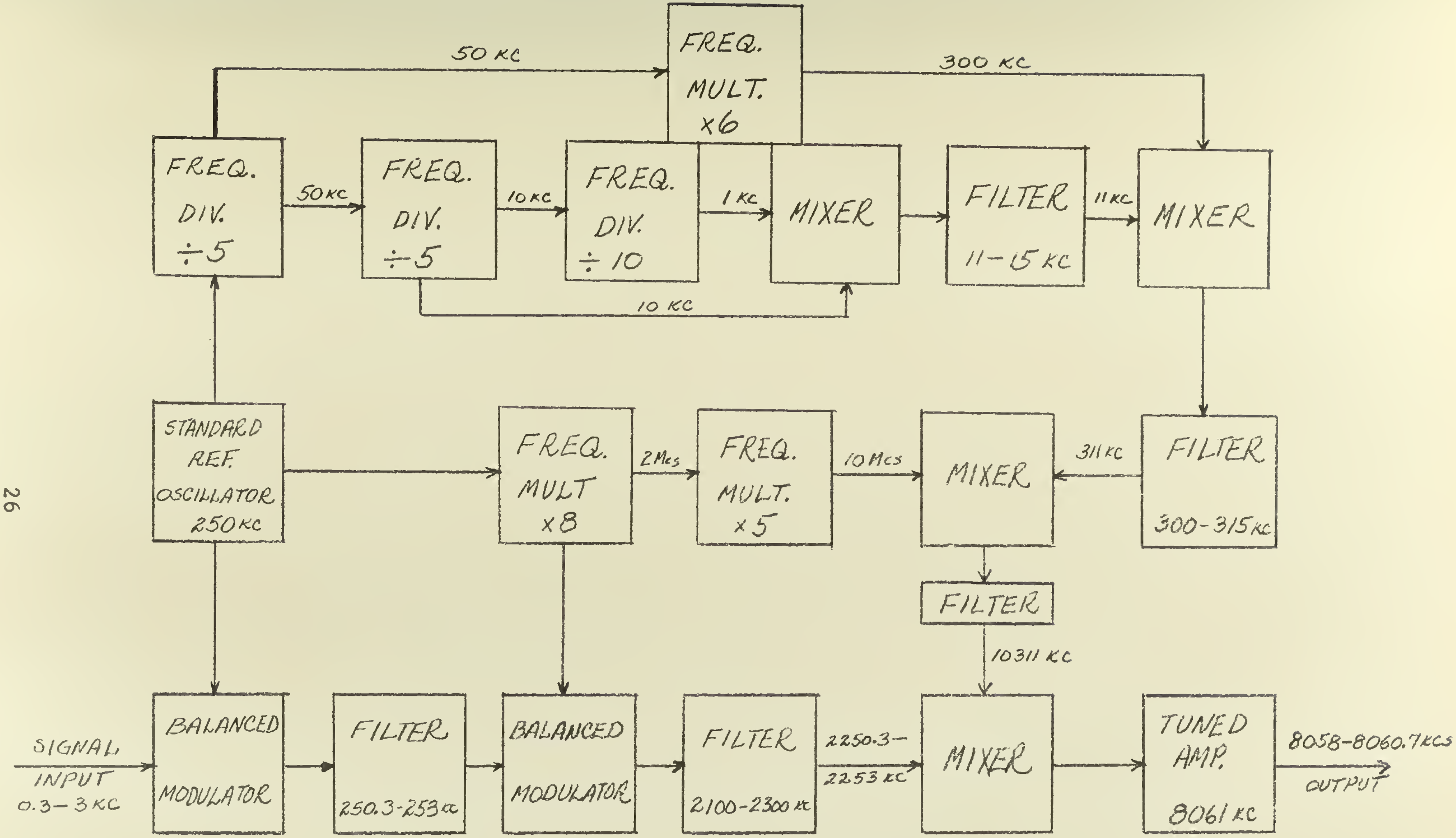
[1] W. L. Firestone, SSB Performance as a Function of Carrier Strength, Proc. I.R.E. 44, pp. 1839-1848, Dec., 1956.

[2] R. L. Craiglow and E. L. Martin, Frequency Control Techniques for S.S.B., Proc. I.R.E. 44, p. 1702, Dec., 1956.

entirely by the aging of the crystal resonator.

The single crystal frequency synthesis system involves the use of frequency dividers, multipliers, and mixers to obtain the desired output from an initial reference input of from 100 to 300 kilocycles. Figure 6 shows a simplified example of a frequency synthesis system forming an output channel of 8061 kcs. The output of each mixer is filtered to obtain the desired modulation product and each mixer is operated as close to its parabolic curvature characteristics as possible in order to avoid unwanted intermodulation products in the range of the bandpass filters. For example, if the mixer forming the 11 kc product were operated so that the power series representation of its dynamic transfer characteristic contained large third and fourth power coefficients, the output of this mixer would consist of 6-14 kcs signals in 1 kc steps, not to mention the harmonics of the 10 kcs input. The filter shown would then pass the 11, 12, 13, and 14 kcs outputs, forming four channels instead of one. The fact is that the major difficulty encountered in the design of a frequency synthesizer is the presence of these unwanted intermodulation products generated in the combining mixers. Extensive filtering and extremely careful selection of combining frequencies are required for even the simplest of circuits. For a 2-30 Mcs transmitter with channels 1 kc apart, 28000 channels would be required. For a frequency synthesis system to operate on 28000 channels effectively, the complexity would be astounding as would be the cost of such a system.

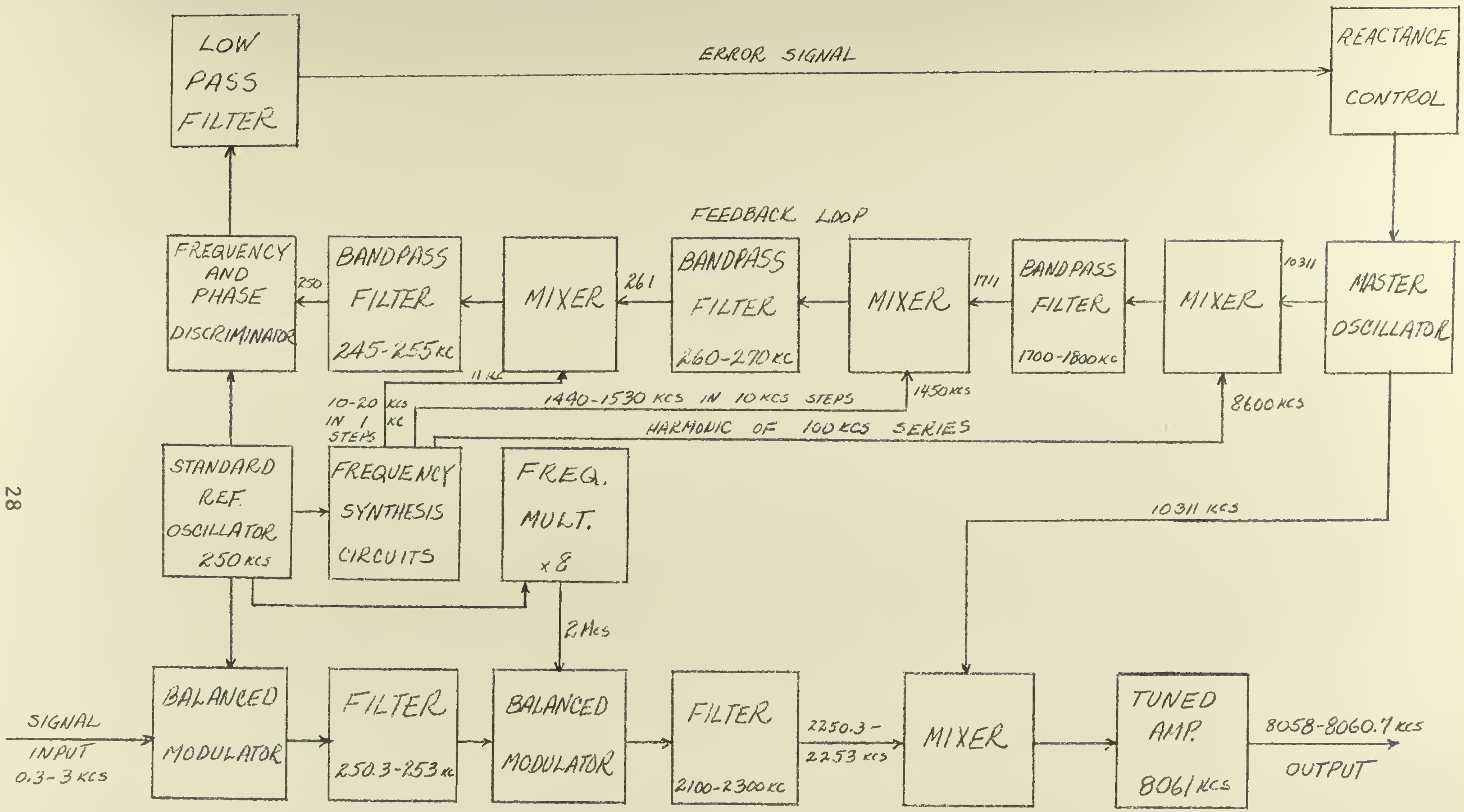
It is possible to retain the advantage of the frequency synthesizer and avoid many of its spurious frequency problems by use of a synthesizer to provide a reference signal to control the frequency of



FREQUENCY SYNTHESIS OF 8061 kcs

FIGURE 6

a variable frequency oscillator. A circuit of this kind is known as a stabilized master oscillator and is shown in Figure 7. The output channel of 8061 kcs is obtained with reference oscillator accuracy by use of a servo loop to feedback error signals derived by comparison of the master oscillator frequency and phase with that of the reference oscillator. A frequency synthesis network is employed for the purpose of accurately presenting the master oscillator output to the discriminator at the reference oscillator frequency. For example, when the master oscillator is to be set up on a certain frequency, the mixers in the feedback loop are fed selected frequencies from the synthesizer in order to heterodyne this output frequency down to 250 kcs. If the master oscillator, in the example shown, were initially set up on 10315 kcs, the bandpass filters would pass 1715 kcs, 265 kcs, and 254 kcs respectively. The frequency discriminator would then send a signal to the reactance control which would react to bring the master oscillator frequency down. When the oscillator reached 10311 kcs, no error signal would be sent out and the oscillator would stabilize at that point. The use of a phase error signal in the control loop insures that the master oscillator error will be in terms of degrees of phase angle rather than cycles. A frequency discriminator is still needed, though, to initially position the master oscillator for the pull in range of the phase discriminator is not usually large enough to handle the initial tuning error of the master oscillator. The reactance control shown could be a saturable inductor, reactance tube, or other like mechanism. The complexity of the synthesizer is much less than that of the pure frequency synthesis system due to the fewer number of selected frequencies required to make 28000 channels. Although the stabilized master oscillator avoids many



STABILIZED MASTER OSCILLATOR OUTPUT OF 8061 kcs

FIGURE 7

of the unwanted intermodulation products of the standard synthesizer, care must still be taken in determining the frequencies to use in the mixers, for it is possible to phase modulate the master oscillator with unwanted products.

To obtain the required stability in a single sideband system, it is necessary to use some method of frequency synthesis with all frequencies being derived from a standard reference oscillator. If just a few well spaced channels were required, the conventional frequency synthesis system would be more than adequate and surely more reliable, but with the requirement for 28000 channels throughout the 2-30 Mcs range, the stabilized master oscillator can afford greater simplicity and greater rejection of spurious intermodulation products. For the 2-30 Mcs transmitter under study, the stabilized master oscillator is definitely the more advisable.

CHAPTER V

THE AMPLIFIER

In order to realize the full advantages of single sideband transmission, it is necessary that the distortion products transmitted should be of extremely low magnitude, preferably 50 db or more below sideband output. These spurious products, which spill over into the adjacent channels, must not be allowed to interfere with the intelligibility of adjacent channel signals under low signal to noise conditions. The main requirements for the amplifier stages of a single sideband transmitter are, therefore, low distortion and reliability. Considering an airborne transmitter, the additional requirements of efficiency of weight, space, and power consumption are of importance. To minimize distortion it would be desirable to use class A_1 amplification, but plate dissipation restrictions of high voltage transmitting tubes make this an impossibility. Today two types of amplification stages are in use: linear class AB and the Kahn method of envelope elimination and restoration. These two will be discussed below with a view to satisfying the above requirements for an airborne single sideband transmitter.

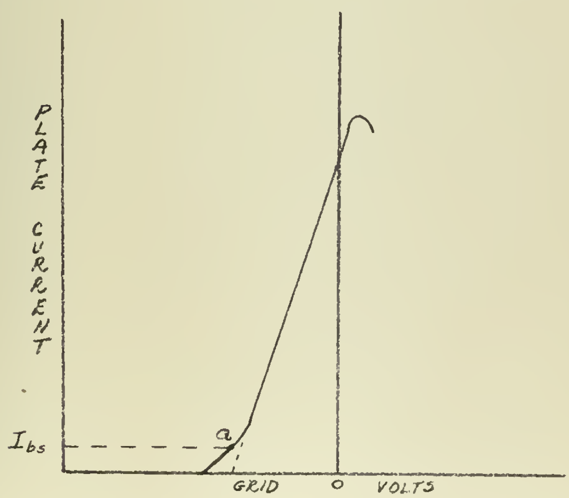
First, it is well to consider the distortion producing mechanism and the distortion products that are most bothersome. As discussed previously, the dynamic transfer characteristic of any tube is non-linear and can be represented by a power series of the form:

$$i_p = a_0 + a_1 e_g + a_2 e_g^2 + a_3 e_g^3 + a_4 e_g^4 + \ldots\ldots.$$

where i_p is the instantaneous plate current and e_g is the input grid voltage signal. This input signal, if voice, is made up of a complex wave that may cover a frequency spectrum of from 2,000,300 cps to

2,003,000 cps if an upper sideband signal of a 2 Mcs suppressed carrier. If we consider just two of these components, in the lower end of the spectrum where the more intense components are located, we can get a fair idea of the distortion problems involved when all components are considered. Considering the 2,000,300 cps and the 2,001,000 cps components, the linear term of the power series would pass these signals as such. The second order term would create the sum and difference frequencies, 4,001,300 cps and 700 cps, both of which would be well out of the pass band. The third order term would create $2F_1-F_2$ and $2F_2-F_1$ products, 1,999,600 cps and 2,001,700 cps, which would surely be bothersome in this and the adjacent lower channel. In like manner it can be ascertained that all even order terms create intermodulation products well out of the pass band, and all odd terms create bothersome products; $3F_1-2F_2$ and $3F_2-2F_1$ for the fifth order, and $4F_1-3F_2$ and $4F_2-3F_1$ for the seventh.

The ideal amplifier would be one, then, whose non-linearity was confined to the second order only. Unfortunately, the dynamic transfer characteristics of tubes are not true parabolas, but are, rather, a combination of approximately linear characteristics with a curved portion near the cut-off region. See Figure 8.



DYNAMIC TRANSFER CHARACTERISTIC

FIGURE 8

It has been found[1] that by choosing an operating point close to the center of the curved portion of the characteristic, nearly parabolic operation in this region is obtained. Operating at this ideal point, in a combination of the parabolic and linear regions, gives no worse than second order distortion products of any bothersome magnitude. At this ideal operating point, however, the static plate current, I_{bs}, turns out to be just the right amount to insure maximum plate dissipation in high voltage tubes with no signal whatsoever applied to the grid. Obviously, some compromise must be established. In practice the compromise is one of operating with a lower than ideal screen voltage and driving the grid positive. The grid is then loaded heavily with a fixed resistor in order to minimize non-linear grid characteristics. This compromise enables high gain tubes to be employed with distortion 30-35 db below signal output. For high power transmitters several stages of amplification must be employed, and the distortion products per stage are added approximately vectorally to obtain total distortion output. High gain tubes are employed in order to reduce the number of stages, but even then some means of distortion reduction is necessary if the signal to distortion ratio is to be a tolerable 35-40 db, much less the 50 db desired. W. B. Bruene of the Collins Radio Company has done work on this problem[2] and has ascertained that 10 db of r.f. feedback across three stages produces 10 db distortion reduction, giving a signal to distortion output of 35-40 db. A second method of distortion

[1] W. B. Bruene, Linear Power Amplifier Design, Proc. I.R.E. 44, pp. 1755-1756, Dec., 1956.

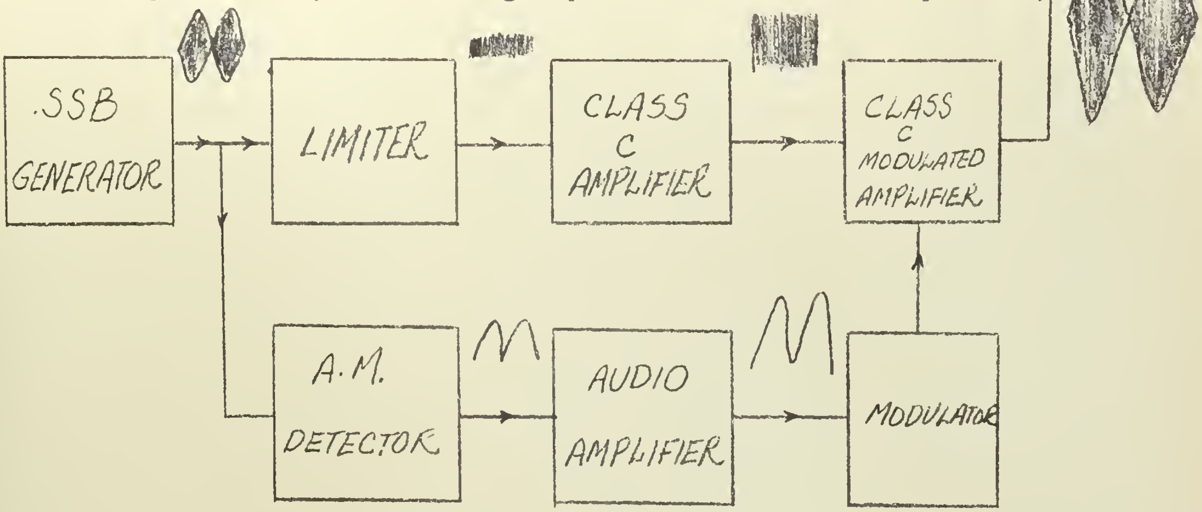
[2] W. B. Bruene, Distortion Reducing Means for Single-Sideband Transmitters, Proc. I.R.E. 44, pp. 1760-1765, Dec., 1956.

reduction investigated by Mr. Bruene was that of envelope distortion cancelling modulation which employs a form of amplitude modulation to restore the shape of the distorted single sideband envelope. This second method requires complex design and intricate balance, and its capabilities have been quite limited up to the present time. Mr. Bruene was able to obtain 50 db signal to distortion outputs in the laboratory using a combination of the above two methods, but he has indicated reason to doubt the effectiveness of the envelope modulation due to the unpredictability of the non-linear element characteristics.

As can be seen, the difficulties encountered in linear amplification are multifold. In 1952 L. R. Kahn presented a method of amplification involving envelope elimination and restoration.[1] This method employs class C amplifiers and high level modulation that makes the distortion independent of the number of stages and insures greater efficiency of power consumption, especially under zero signal conditions. Figure 9 shows a simplified block diagram of the Kahn method, waveforms shown being the result of modulating the single sideband generator with two equal amplitude tones of different audio frequencies. The single sideband signal from the generator can be shown to be made up of two parts, one, amplitude modulated and the other, phase modulated. The limiter shown passes only the phase modulated part which is then amplified by a class C amplifier and drives the final modulated power amplifier. The non-linearities of a class C amplifier do not introduce distortion in a phase modulated

[1] L. R. Kahn, Single-Sideband Transmission by Envelope Elimination and Restoration, Proc. I.R.E. 40, pp. 803-806, July, 1952.

signal. An amplitude modulation linear detector passes the amplitude modulated part which, after being amplified in an audio amplifier,



ENVELOPE ELIMINATION AND RESTORATION
FIGURE 9

modulates the phase modulated signal in the class C power amplifier. This system has been used in practice with a high quality modulator and has produced a 400 kw peak output with a signal to distortion ratio of 32-35 db.[1] While simple in appearance, the use of this system entails critical adjustments, for the phase and amplitude relations of the two components must be preserved. This means that the time delay through the two paths must be the same. Inherently, the time delay for the phase modulated branch is negligible, whereas the audio branch has a considerable time delay by comparison. A time delay circuit must be added to the phase modulated branch which should insure constant time delay throughout the frequency range of the transmitter. Also, the average direct current level of the generated signal must be ascertained and used to control the operating point of the final amplifier in order

[1] L. R. Kahn, Comparison of Linear Single-Sideband Transmitters with Envelope Elimination and Restoration Single-Sideband Transmitters, Proc. I.R.E. 44, pp. 1706-1712, Dec., 1956.

to insure proper amplitude relations.

Using high level modulation, the size and weight of the modulation equipment would surely be a relative disadvantage in an airborne transmitter. Also, the critical requirements for phase and amplitude relations cast some doubt on the relative reliability of this method in comparison with linear amplification. Mr. Kahn shows that the operating efficiency of this system is 69 percent compared to 47.1 percent for linear amplifiers, but even of more importance is the saving in plate power under no signal conditions. There is no doubt that Mr. Kahn's method will have many applications, but for an airborne transmitter, the author does not believe that the saving in power consumption is worth the increase in size, weight, and complexity. Considering the present state of development, the use of linear amplifiers with r.f. feedback best satisfies the requirements for an airborne single sideband transmitter.

CHAPTER VI

CONCLUSIONS

At the present state of development of the single sideband system of radio communication, there are many methods proposed as to how to most efficiently transmit the required signal. These various methods have been discussed in the preceding chapters with a view to obtaining the optimum system for an airborne transmitter.

The basic requirements for an airborne transmitter, listed in the Introduction, can be interpreted, for the single sideband transmitter, to require additionally:

1. Preferably 40 db or more signal to distortion ratio, so that the distortion products of a strong signal will not over-ride an adjacent channel weak signal.

2. Simplicity of design in order to ease the maintenance problems.

3. Use of passive elements where possible in order to maintain stability in critically adjusted circuits.

4. Employment of techniques to minimize size, weight, and power requirements for the desired transmitter.

The optimum system for an airborne transmitter was deemed to include:

1. Balanced modulators employing varistors due to their stability and essentially infinite life. A power level of from 0.1 to 0.2 watts is all that is necessary to drive the amplifiers; therefore, higher power generation is not required.

2. The filter method of signal generation due to the fact that for a high signal to distortion ratio output from a variable frequency transmitter, the filter method is the more simple and easy to maintain.

The phase shift method would require a greater number of elements and is much more sensitive to slight changes in parameters than the filter method. The Weaver method shows much promise due to the fact that adjacent sideband interference is eliminated, but the author imagines that amplitude, phase, and time delay requirements are just as critical, if not as difficult to maintain, as for the phase shift method.

3. The stabilized master oscillator for frequency stability. The single crystal frequency synthesis system would be ideal for fewer, more widely spaced channels, but would be too complex to use for 28000 channels.

4. Linear amplification because it offers the best compromise in size, weight, and power requirements in the 500 to 1000 watt class. The Kahn method would be ideal for conversion of existing amplitude modulation transmitters to single sideband and for higher power applications where power savings and signal to distortion advantage would be pronounced.

The system chosen is not, by any means, the ultimate, but at the present state of developments it seems to be the optimum to meet the above requirements. The Collins Radio Company has built for the Air Force the AN/ARC-58 which is now under development at the Rome Air Development Center. It is of interest to compare the vital statistics of the transmitter of this airborne single sideband transceiver with the ART-13 transmitter it will replace:

	ART-13	AN/ARC-58
Peak power output	90 watts	1000 watts
Peak sideband power output	15 watts	1000 watts
Power input	1036 watts	1425 watts
Weight including power supply	86 pounds	64.5 pounds

The development of this single sideband transceiver has and should continue to further the applications of this system of radio communication.

CHAPTER VII

EXPERIMENTAL RESULTS

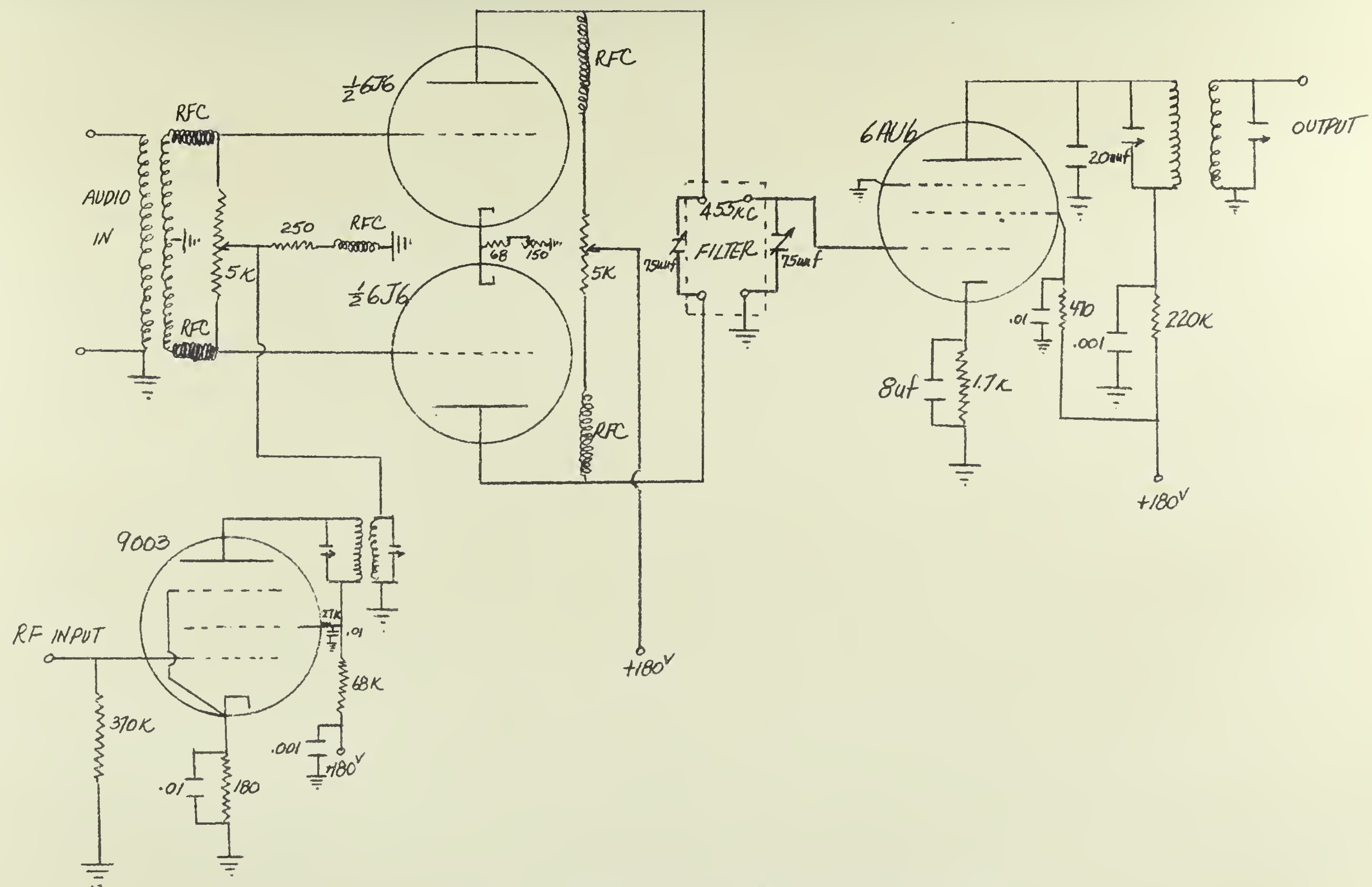
As stated in Chapter II, laboratory research included building and testing balanced modulators employing tubes, crystal diodes, and transistors. The objective of this chapter is to describe the modulators in regard to circuitry, balancing mechanisms, degree of balance, stability of balance, and intermodulation distortion.

EQUIPMENT

All three circuits were built sequentially in the same chassis which contained a r.f. buffer amplifier, a 455 kc, 3000 cycle bandpass Collins mechanical filter, and a linear amplifier. See Figure 10. The output load consisted of a Tektronix 511A oscilloscope and a Hewlett-Packard 400A vacuum tube voltmeter. A 400A vacuum tube voltmeter was connected to the input side of each modulator for obtaining the correct amplitude of carrier and signal voltages. Radio frequencies varying from 405 through 505 kcs were supplied to the buffer amplifier by an LP-5 signal generator. To obtain good sideband separation, an audio signal of 10 kc was supplied to the audio input transformer by a Hewlett-Packard 200B audio oscillator.

PROCEDURE

Each run recorded consisted of choosing specific amplitudes of carrier and signal voltages and measuring the sideband and intermodulation products out to the fifth order on each side of the suppressed carrier. Sidebands and distortion products were measured on the output vacuum tube voltmeter by varying the r.f. carrier from 405 - 505 kcs and obtaining the desired quantity through the sharp bandpass 455 kcs filter. For



CHASSIS LAYOUT FOR BALANCED MODULATOR TEST

FIGURE 10

example, the lower sideband was obtained on each run with an input of 465 kcs, and the upper fifth order distortion at 405 kcs r.f. input. The r.f. double tuned circuit used in the buffer amplifier plate circuit was peaked each time the frequency was changed to insure that the carrier input amplitude did not change during the run.

VACUUM TUBE MODULATOR

Shown in Figure 10, the vacuum tube modulator employed two balancing mechanisms, one in the plate and one in the grid circuit. The grid balance was the more effective of the two, although both were necessary for optimum balance. Another variable built into this circuit was a 150 ohm grid bias potentiometer which varied the operating point on the dynamic transfer characteristic so as to investigate the effect of the curvature of this characteristic.

To completely balance the carrier in any balanced modulator, the power series coefficients of each modulator must be identical, the input signals must be balanced, and the phase lag through each path must be the same. In practice this is impossible, but preliminary balance can be obtained by matching, as nearly as possible, signal paths and modulator characteristics. Inherent unbalance is then removed by varying the grid signal balance and the plate load to alter one modulator characteristic with respect to the other. In the vacuum tube modulator it was possible to find the point of maximum parabolic curvature, but even this point showed other power series coefficients as witnessed by the intermodulation products. Eleven recorded runs were made with this modulator, four of which were made at the best possible operating point which was at an E_c of minus three volts.

Resume of results:

E_c in volts	Carrier Input in volts	Carrier Signal	Sideband in volts	Carrier	3rd	5th
				in db below sideband		
-1.7	0.95	1.08	1.8	-36	-14	-32.5
-2.1	1.04	0.98	3.4	-40	-21	-40*
-2.1	1.04	3.15	1.17	-30.5	-40*	-40*
-2.1	1.04	16.25	0.3	-19	-40*	-40*
-2.5	1.11	0.74	5.2	-43.5	-22	-40*
-2.5	1.08	1.89	2.6	-39	-40*	-40*
-2.5	1.07	3.32	1.0	-28.2	-40*	-40*
-3.0	1.14	0.63	5.6	-49	-26.2	-40*
-3.0	1.11	2.02	2.1	-35.1	-40*	-40*
-3.0	1.17	4.77	0.9	-40.8	-40*	-40*
-3.0	0.25	1.00	0.3	-27.5	-40*	-40*

* greater than -40 db and unreadable.

Results show the advantageous effect of the more negative grid bias as far as intermodulation distortion with respect to sideband amplitude is concerned. Also shown is the effect on intermodulation distortion when the input signal amplitude is greater than the carrier. Best results over-all were obtained when the carrier was two to five times greater than the signal. The vacuum tube modulator was easy to balance but the stability of balance left much to be desired. Each time any small change was made to either carrier or signal input, the modulator required rebalancing. No doubt that better matched tubes would have been more stable, but even then, with the inevitable aging, unbalance would occur.

CRYSTAL DIODE MODULATOR

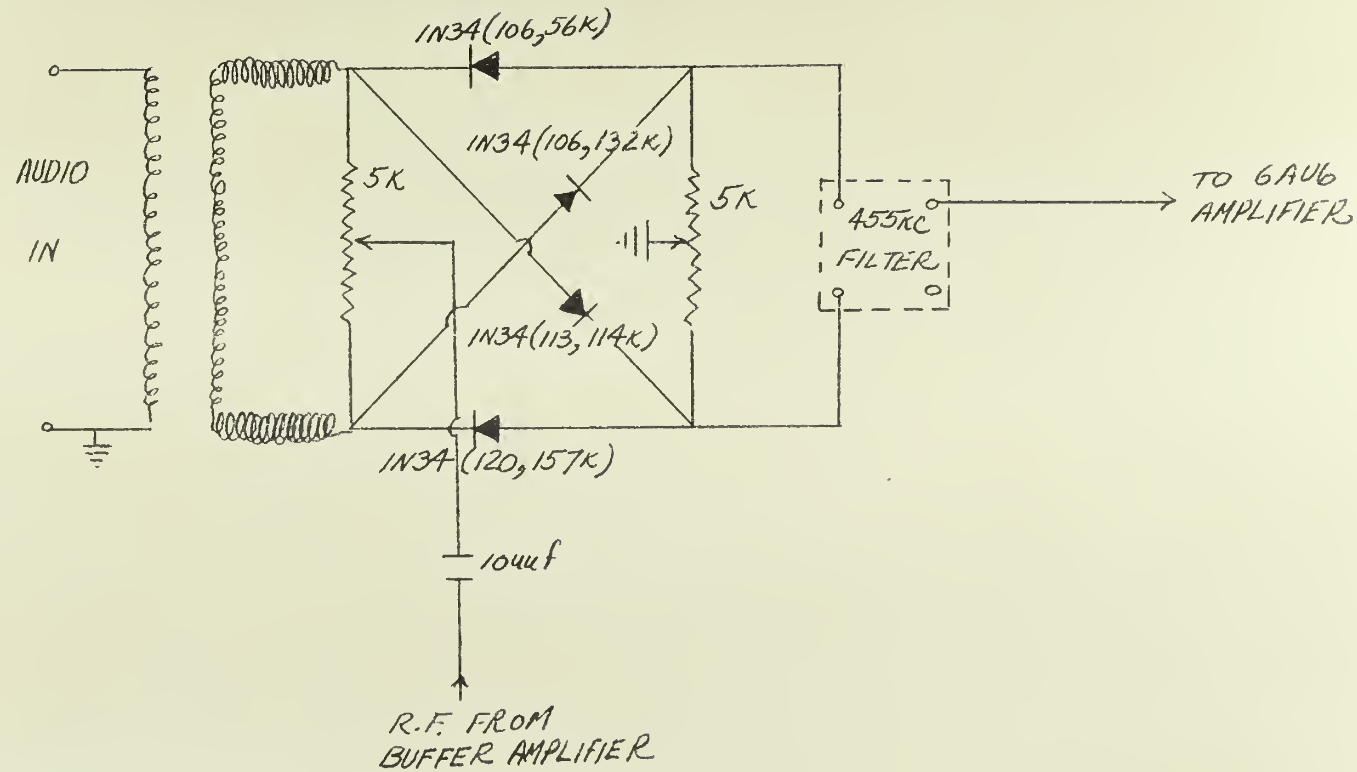
A double balanced ring modulator employing 1N34 crystal diodes, as shown in Figure 11, was investigated in this project. Care was taken to match the resistance of the two paths as far as possible, the forward resistance of one path being 226 ohms to 219 ohms of the other. The respective reverse resistances, while not as critical, were 213 K to 246 K. Five different runs were recorded varying the ratio of carrier to signal. Again two balancing mechanisms were employed, one to vary the carrier input to each leg slightly and the other to vary the balanced load. Both mechanisms were useful in obtaining balance, the ease and degree of which compared favorably with the vacuum tube modulator.

Resume of Results:

Carrier Input in volts	Carrier / Signal	Sideband in volts	Carrier (in db below sideband)	3rd (in db below sideband)	5th (in db below sideband)
1.1	1.53	1.25	-34	-17	-32
1.1	2.45	0.79	-30	-17	-32
1.2	5.85	0.70	-29	-17	-29
1.2	10.4	0.44	-26	-27	-40*
1.2	20.0	0.19	-17	-40*	-40*

* greater than -40 db and unreadable.

The results show the reduction of intermodulation products as the carrier to signal ratio is increased substantially. Carrier balance remained practically the same for all runs, the disadvantage showing up in the latter runs being in the sideband amplitude. Balance had to be changed very little between runs but more between the first four runs than between the latter two. Aging being no problem here, the stability of



DOUBLE BALANCED RING MODULATOR USING 1N34's

FIGURE 11

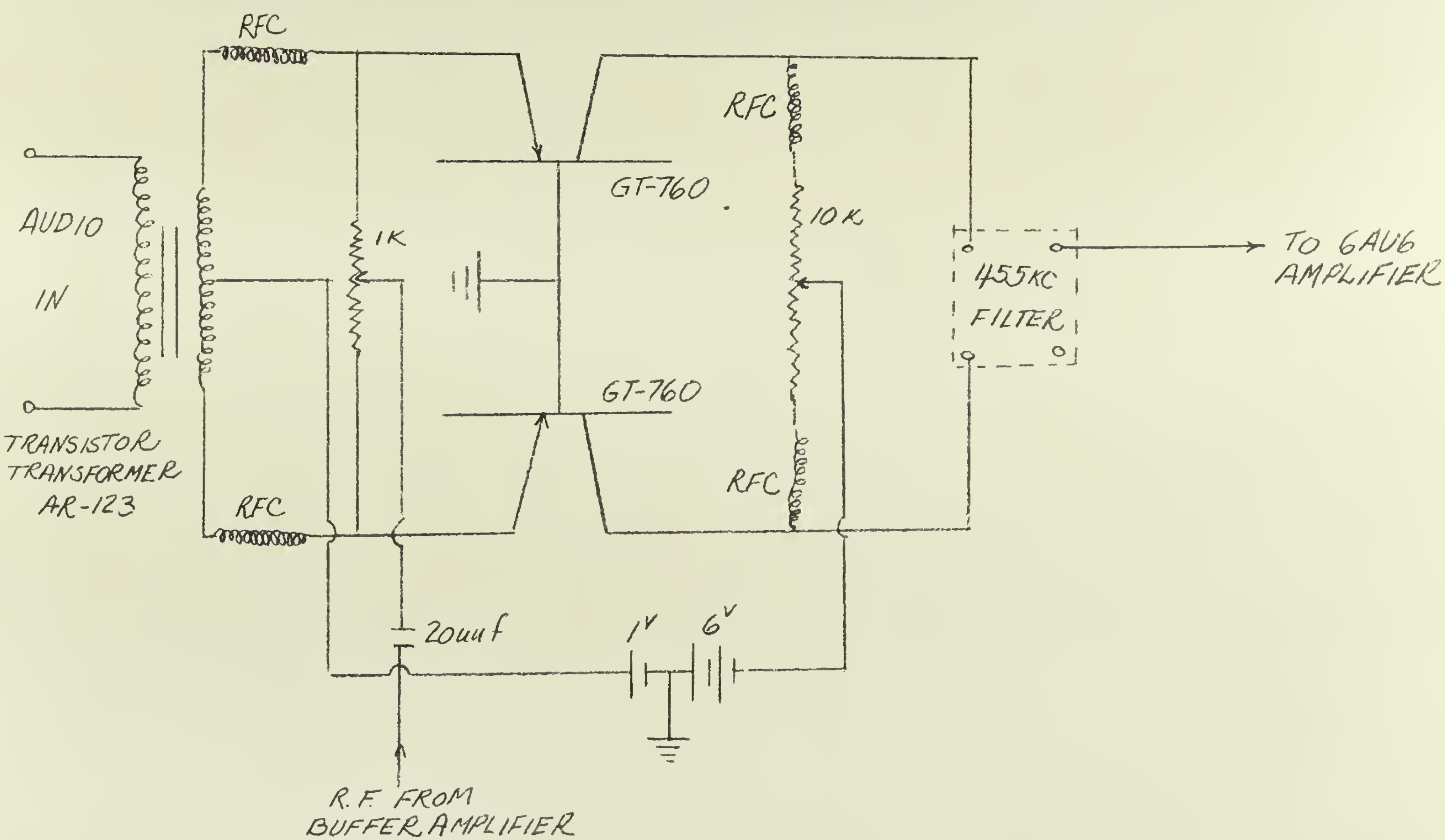
balance could be maintained for a long period of time at high carrier to signal ratios.

TRANSISTOR MODULATOR

Shown in Figure 12, the transistor modulator employed two GT-760 P-N-P transistors in the grounded base configuration. Prior to building the modulator, one of the transistors was checked in an experimental chassis in order to ascertain the optimum amounts of bias and the best configuration to obtain maximum second harmonic output. The grounded base configuration was chosen with an emitter bias of plus 4.8 volts and a collector bias of minus five volts. Upon building the modulator, it was found difficult to obtain the correct bias without breaking down the high collector to base resistance. Due to this fact, it was necessary to use slightly more negative collector bias and less positive emitter bias than desired. The balancing mechanism consisted again of a method of varying carrier strength on the two emitters and a variable load balance. The load balance had practically no effect, but the carrier balance was quite effective. The carrier balance was 10-20 db poorer than for either of the other two modulators, undoubtedly due to a greater mismatch between the two transistors than between the characteristics of the other components tested.

Resume of results:

Carrier Input in volts	Carrier Signal	Sideband in volts	Carrier	3rd	5th
			in db below sideband		
0.65	0.412	1.8	-17	-9	-13
0.012	0.209	0.9	-18	-18	-33
0.0015	0.0555	0.09	-1	unreadable	



BALANCED MODULATOR USING GT-760 P-N-P 455 kc TRANSISTORS

FIGURE 12

The runs were made sequentially with less carrier and less carrier to signal ratio. It was necessary to make the carrier to signal ratio low in order to obtain any sideband output whatsoever. The greatest obstacle seemed to be obtaining the correct bias. If this had been obtained, I believe that a greater carrier to signal ratio would have given better results. I also believe that the transistors should be matched as nearly as practicable in order to obtain effective carrier balance. The stability of balance was quite poor, for any small change in either input necessitated a complete rebalance of the carrier. Two-tone tests[1] were run on all three modulators at each one's optimum operating point. No significant difference in linearity was apparent upon examination of photographs of the three tests, but the transistor modulator showed itself to be by far the most sensitive to signal overload.

I do not believe that transistors have application in the lower frequency modulators of a single sideband system, but possibly, if great care were taken in design, could have application in the r.f. balanced modulator that feeds the linear amplifiers. Considering the simplicity of design, lack of power requirement, bilateral ability, and stability of the crystal diode modulator, I believe this to be the best modulator for airborne use.

[1] P. J. Icenbice, Jr. and H. E. Fellhauer, Linearity Testing Techniques for Sideband Equipment, Proc. I.R.E. 44, pp. 1775-1781, Dec., 1956.

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